A.R.F. PRODUCTS, INC.

ENGINEERS AND MANUFACTURERS OF PRECISION ELECTRONIC EQUIPMENT

Contains System Equations

MONTHLY TECHNICAL
LETTER REPORT
NUMBER NINE
PERIOD ENDING
30 SEPT. 1980

MINE WIRELESS COMMUNICATION SYSTEM

HO 308004

A.R.F. PRODUCTS, INC.

Raton, New Mexico

Boulder, Colorado

TITLE   WIRELESS MF COMMUNICATION SYSTEM

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MONTHLY TECHNICAL LETTER REPORT
NUMBER NINE
PERIOD ENDING 30 SEPT. 1980

CONTRACT NO. HO 308004

DEVELOPMENT OF DEMONSTRATION
HARDWARE FOR MEDIUM FREQUENCY
MINE WIRELESS COMMUNICATION SYSTEM

PREPARED FOR

UNITED STATES DEPARTMENT OF THE INTERIOR
BUREAU OF MINES
4800 FORBES AVENUE
PITTSBURGH, PA 15213

BY: _____
Larry G. Stolarczyk, Sc.D.
Director of Research

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EXECUTIVE SUMMARY

STATUS REPORT

Mining Research
Mine Health and Safety Research
      for
Period Ending 30 September 1980

PROGRAM Wireless MF Communication System

DATE STARTED January 1980

TPO (USBM) Robert I. Chufo (P-SRC)

CO (USBM) Alan C. Bolton, Jr. (Mr. Patrick Neary)

RESEARCH AREA Industrial Type Hazards

SCHEDULED COMPLETION   Phase I   29 September 1980
                       Phase II   29 June 1981
                       Phase III   29 April 1982

PROJECT OBJECTIVE Development of demonstration hardware for
medium frequency mine wireless communication system.

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PROGRESS AND STATUS (PHASE I)

The signal levels appearing at the receiver (A1) input terminal have been determined for the following signal transmission mediums:

Transmission medium	Transmission mode
Trolley wire	Transmission line (trolley wire to track separation of 3 meters)
AC power distribution cable	Transmission line (conductor separation of 0.04 meter)
Telephone lines	Transmission line (conductor separation of 0.04 meter)
Cellular	Coal seam mode

The signal levels have been determined for loop to loop via the line (LLL) communications, as well as for LLL communications links that use passageway repeaters. Passageway repeaters are directly connected to the transmission line to enhance the calling range by significantly increasing the signal current on the transmission line.

The cellular communication system signal levels have been determined by using the measured field strength data from the U.S.B.M. Collins (HO 366028) and ADL (HO 340045) reports.

Bear in mind that the signal level estimations have been made on the basis of a well behaved transmission line structure. In practice, insertion and missmatch loss will degrade the expectation.

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This report adds to the current body of underground communication system design knowledge by:

* accurately defining (measuring) the 12 dB SINAD sensitivity of the system receivers and then stating the equivalent magnetic field strength (Hs) in uA/m;

* showing that when a passageway repeater is used in loop to loop via the line communication link that the received signal level dependence on line current is only the _first_ power of distance; and

* maximizing the transmit magnetic moment (M) is extremely important in the design of the transmit antenna network.

Since the 12 dB SINAD sensitivity of the receiver (A1) is known to be one (1) microvolt, the magnetic field (Hs) required to produce the 12 dB SINAD signal can be accurately determined   Thus the Hs data can be used to correct the Collins seam mode range expectation. At 600 kHz, Hs has a value of -20 dB re 1 uA/m, and degrades to -1 dB re 1uA/m at 60 kHz.

The received signal level (V_L) in a loop to loop via the line communication link has a _third_ power dependence on distance (R) as seen in the following mathematical representation:

$$V_L \propto \frac{f^2}{R^3} M \, b \, A_R N_R$$

where f is the operating frequency, M is the transmit magnetic moment,

b is the separation distance of conductors in the transmission line,

and $A_R N_R$ are the receiving loop antenna design parameters.

The signal level (V_L) dependence on line current and distance (R) is given by

$$V_L \propto \frac{f}{R} A_R N_R \, I_{LINE} \cdot$$

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Since a passageway repeater can greatly increase line current, and the distant receiver signal level (V_L) has only a first **power** dependence on distance (R), the "called" transceiver path distance is greatly increased.

The repeater will increase the "called" transceiver distance from the transmission line by a factor of 2^N over the LLL communication technique!

The passageway repeater also increases the "calling" transceiver distance from the transmission line. The repeater receiver signal level has a **second** power dependence on distance. Thus the "calling" range will not show the dramatic improvement as seen in the "called" range.

The signal level data presented in this report has been **normalized** by a unit magnetic moment (1 ampere - meter^2). Simply multiply the signal level data by the actual transmit moment (M) to obtain (denormalize) the actual signal level estimation.

The normalization method focuses attention on the importance of the magnetic moment (M). A strong factor in the optimization of system performance is the maximization of M for a given transmitter output power. The dependence of magnetic moment (M) on transmit power (P_0) and the series resistance (R_L) of the tuned loop antenna ($R_L = \omega L/Q_{CKT}$) is given by

$$M = N A \sqrt{\frac{P_0}{R_L}}$$

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It is clear from the above equation that increasing power from 1 watt to 4 watts only increases the magnetic moment by a factor of two with R_L held constant. M can be increased at any transmit power level by decreasing R_L. However, this optimization approach is difficult because Q_{CKT} is constrained by the occupied bandwidth of the narrow band FM signal.

Single tuned loop antenna structures cannot be optimized to increase magnetic moment above 1.6 ampere - meter^2 (P_O = 4 watts). However, a multiple pole-zero loop antenna can be used to increase M above 1.6.

There exists a concern that MF signals in the vicinity of a blasting area will create hazardous situations. The induced current in blasting cable is estimated to be a maximum of 500 microamperes.

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1.0   INDUCED TROLLEY WIRE SIGNALS

The trolley wire and rail will be used as the communication media in many underground mines.

In this section, the signal level induced on the trolley wire and in the input circuits of a distant receiver by a first man-pack transmitter will be determined.

The work of Albert A. Smith Jr. and the U.S.B.M. Arthur D. Little, Inc. report (Contract No. HO 346045 -- Section IV) will be used to determine the signal levels.

The trolley wire signal transmission medium (two wire line approximation) will be excited by a circular loop with the geometrical alignment shown in Figure 1.0.
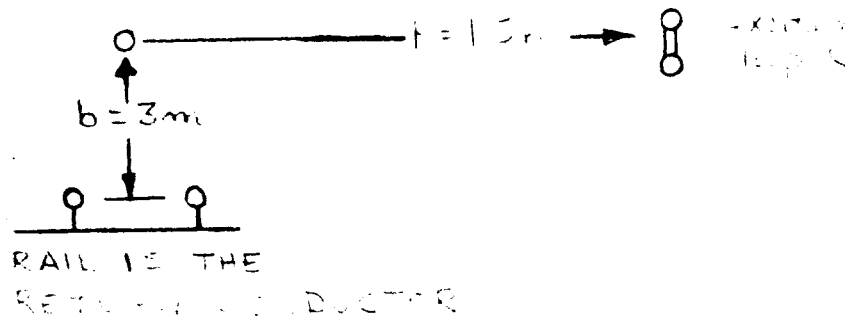

Figure 1.0   The Aligned Geometry

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The physical distance between the trolley wire and track is b in meters

and the distance between the loop and the line is R in meters.

The induced voltage source (emf) in the two wire line is given by

$$V = \frac{\mu_o \, fMb}{R^2} \qquad 2\pi R \ll \lambda \qquad (1)$$

$$= \frac{(4\pi \cdot 10^{-7})(60 \times 10^{+3}) \cdot 1 \cdot 3}{(1.5)^2} \cong 0.1 \text{ volt}$$

The induced voltage source appears in Figure 1.1.

Figure 1.1 EQUIVALENT TRANSMISSION LINE CIRCUIT
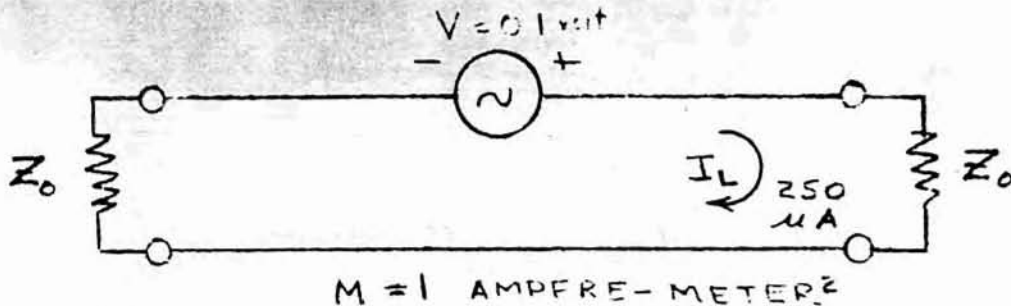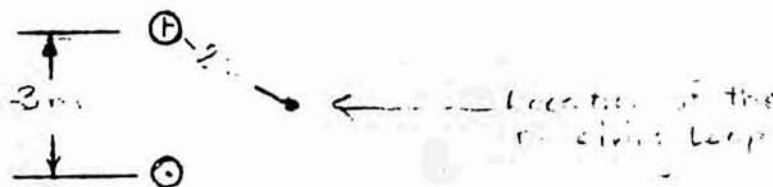FOR INDUCED VOLTAGES AND CURRENTS

The induced transmission line current is given by

$$I_L = \frac{0.1 \text{ volt}}{2(200\pi)} = 250 \text{ uA} \qquad (2)$$

The magnetic field at the location of the receiving loop may be

calculated as the vectorially resultant field (same geometry as the

transmitting side) of contribution of the two conductors. The geometry

is shown below.

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The magnetic field component from each current may be calculated from

$$\frac{I_L}{2\pi(R)} \quad , \quad \text{then}$$

$$H_c = 1.41 \; \frac{250 \; uA}{2\pi(2.25m)} = 24.93 \; \frac{uA}{m} \tag{3}$$

From Faraday's Law, the voltage source (emf) in the receiving loop antenna is given by

$$U_L = \left|\frac{d\phi}{dt}\right| N = \left|j\omega\mu_0 H\right| A_R N_R \tag{4}$$

where A_R is the area (meters^2) of the receiving loop antenna,

N_R is the number of turns in the receiving loop antenna, and

H is the magnetic field strength (ampere/meter) computed from Equation (3).

Table A shows the value of the tuned loop antenna voltage source (emf) for different operating frequencies for each unit of magnetic moment (M).

Table A
Receiving Antenna Loop Voltage (V_L)

$M = 1$ ampere-meter^2; $b = 3$ meters; $N = 7$ turns; $A = 0.2$ meters^2; $R = 1.5$ meters

Frequency kHz	Antenna Voltage uV	λ meters
60	16.53	5000
300	82.65	1000
600	165.3	500

For completeness, Table B shows the magnetic moment for different transmitter output power levels.

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Table B
Transmitter Magnetic Moment (M)

$N = 7$ turns; BW = 10 kHz; $R_L = 4.7$ ohms

Output Power watts	Transmit Loop Current ampere	Magnetic Moment ampere-meter^2
1	0.46	0.81
4	0.92	1.61
20	2.06	3.61

The signal voltage (V_R) available at receiver input terminals is dependent upon the voltage transfer function

$$T_V = \frac{V_R}{V_L} \tag{5}$$

of the tuned loop antenna.

The tuned loop antenna network is shown in Figure 1.2.

Figure 1.2   TUNED LOOP ANTENNA NETWORK
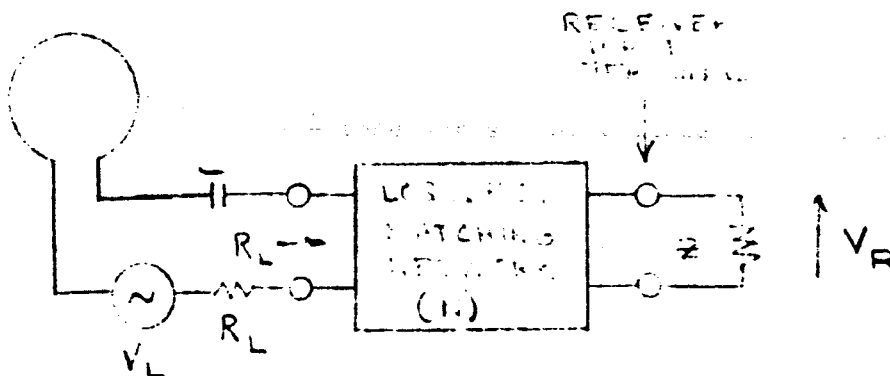
Assuming that the antenna is tuned and matched, the current flowing in the loop antenna is given by

$$i = \frac{V_L \text{ (emf)}^*}{R_L + R_L} \tag{6}$$

where the input impedance of the network (N) is real and equal to the tuned resistance ($R_L = \omega L/Q_{CKT}$) of the loop antenna.

* Bear in mind that maximizing the transmitter loop antenna magnetic moment reduces the value of R_L.

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The signal power delivered to the input of the network (N) is

$$P_{IN} = (i_{LOOP})^2 R_L$$

$$= \left(\frac{V_L}{2R_L}\right)^2 R_L. \tag{7}$$

For a matched lossless network (N), the load plane power (P_R) (seen at the receiver input terminals) is identical to the incident power (P_{IN}) (signal power delivered to the input of the network (N)), hence

$$P_R = P_{IN}.$$

then

$$\frac{(V_R)^2}{Z} = \left(\frac{V_L}{2R_L}\right)^2 R_L \text{ and} \tag{8}$$

$$T = \frac{V_R}{V_L} = \frac{1}{2}\sqrt{\frac{Z}{R_L}} \tag{9}$$

The receiver signal voltage (V_R) can be determined from Equations (4) and (9) as

$$V_R = \left(\frac{1}{2}\sqrt{\frac{Z}{R_L}}\right)V_L \quad \text{or} \tag{10}$$

$$V_R = \left(\frac{1}{2}\sqrt{\frac{Z}{R_L}}\right)(\omega l l_c AN)H. \tag{11}$$

The receiver input signal voltage produced by the signals in Table A are shown in Table C.

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Table C
Receiver Input Signal (V_R)

For each unit of magnetic moment, b = 3 meters

Frequency kHz	Receiver voltage μV
60	27
300	134
600	270

It is interesting to note that the A.R.F. FM receiver achieves a SINAD of 12 dB when the input signal voltage is 1 uV. At 60 kHz the communication system would have a fade margin of 29 dB.

Most of the U.S.B.M. measured data is given in terms of magnetic field strength H. Equation (11) can be used to determine receiver signal levels from the measure data.

The magnetic field required to achieve the 12 dB SINAD sensitivity is given in the table below.

Table D
Magnetic Field Strength (H_S) required for 12 dB SINAD

Frequency kHz	H_S Field μ ampere meter
60	0.920
300	0.184
600	0.092

It should be noted that the Collins report, "Propagation of FM Signals in Underground Mines" (Contract HO 366026), defines the Collins mine wireless radio sensitivity as .013 u A/M at 520 kHz. In view of Table D, the Collins sensitivity speculation is suspect. Appendix C (page 169)

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of the Collins report develops equations identical to Equation (11).

Because the A.R.F. receiver sensitivity (V_{RS}) is already known to be

1 uV for a 12 dB SINAD, the calculation of H_S is straight forward.

In the Collins case, 12 dB SINAD sensitivity had to be calculated.

Apparently an error has been made in their Equation (6).

In summarizing the results of this section, it is evident that voice

quality communications can be established in loop-line-loop signal

paths (see Tables C and D). The following relationship is useful in

estimating changes in loop-line-loop system performance.

$$V_R \propto \frac{f^2}{R^3} \; M \; b \; A_R \; N_R \tag{12}$$

Equation (12) provides the following useful information:

* the distant receiver signal level increases by V_R μV (270 uV at 600 kHz) for each unit increase in transmit magnetic moment (M);

* the transmission line current increases by 250 uA (at 60 kHz) for each unit increase in transmit magnetic moment (M). From Equations (1) and (2) the line current increases with the first power of frequency;

* the signal level (V_R) increases with the second power of the operating frequency;

* the signal level (V_R) increases with the trolley wire track separation distance (b);   and

* the signal level decreases with the third power of the distance (R) between the trolley wire and the loop antennas.

Bear in mind that the above analysis applies to a matched lossless

transmission line.

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1.1 TRANSMISSION LINE CURRENT TO RECEIVING LOOP ANTENNA SIGNAL LEVEL TRANSFER
FUNCTION (PASSAGEWAY REPEATER)

Equation (4) can be used to define the tuned loop antenna voltage (emf)
to line current transfer function as

$$\frac{V_L}{I_{LIN}} \propto \frac{f\, A_R N_R}{R} \quad . \tag{13}$$

The received signal level increases with the __first power__ of frequency
and decreases with the __first power__ of distance (R). Notice that in the
loop-line-loop signal path the signal level (V_R) decreased with the
__third power__ of distance.

A passageway repeater can be used to increase the __calling__ distance from
transmission line to the mobile transceiver. The passageway repeater is
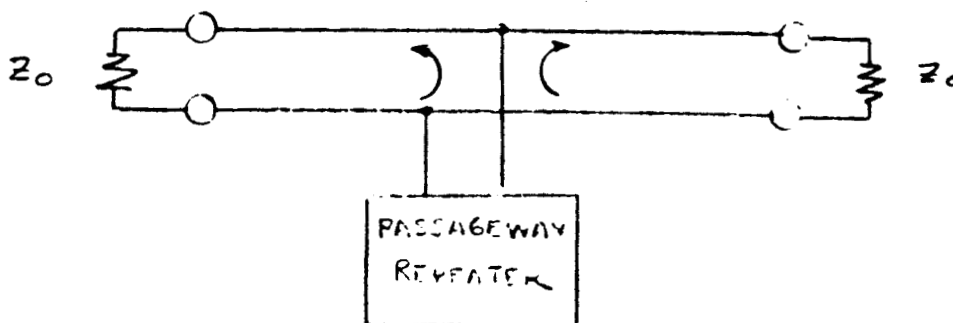shown in Figure 1.3.

Figure 1.3 PASSAGEWAY REPEATER

The passageway repeater will __increase__ the line current to

$$i_L = \sqrt{\frac{20 \text{ watts}}{100 \text{ ohms}}} = .447 \text{ amperes.} \tag{14}$$

Thus the passageway repeater will enable communications over significantly
greater distances from the trolley wire line.

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The passageway repeater increased the line current above the loop-line-loop communication case by

$$\text{System Gain} = \frac{0.447 \text{ ampere}}{250 \text{ u Ampere}}$$

$$= 65 \text{ dB} .$$

Equation (13) shows that the antenna voltage decreases by 6 dB for each doubling of range. The listening range would increase by a factor of 2^{11}.

As a transmission line receiver, the passageway repeater can be modeled as follows:
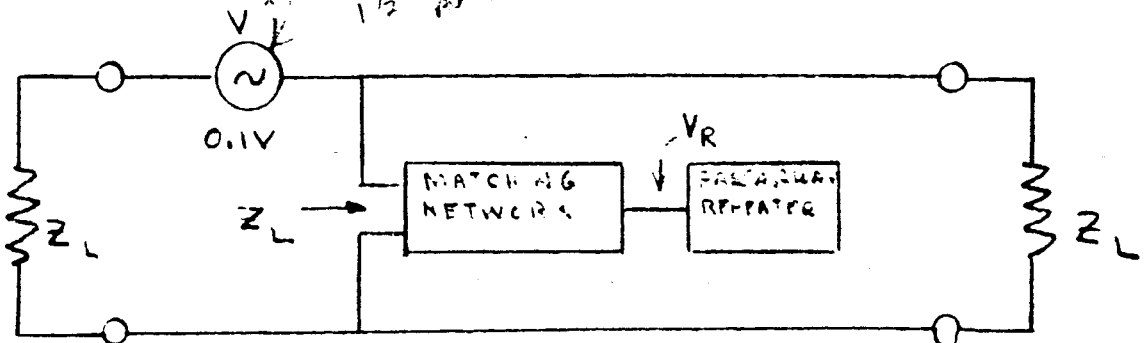
Figure 1.4   PASSAGEWAY REPEATER

The line current can be estimated to be

$$i_L = \frac{.1V}{Z_L + \frac{1}{2}Z_L} = \frac{.1V}{300 \Omega} = 300 \text{ uA}. \tag{15}$$

The received signal level (V_R) is given by

$$V_R = (T) \frac{i_L}{2} Z_L \tag{16}$$

where T is the matching network transfer function.

* Estimated full wave rectification noise (60 V p-p on trolley wire)
  360 Hz signal: 100 kHz, 248 uV; 300 kHz, 27.4 uV; 500 kHz, 9.9 uV.

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At 60 kHz, the received signal level becomes

$$V_R = (\tfrac{1}{2}) \frac{330 \text{ uA}}{2} \cdot (100\Omega)$$

$$= 8.25 \text{ mV} \tag{17}$$

for each unit of magnetic moment (R = 1.5 meters, b = 3 meters).

Equation (1) can be used to estimate changes in system performance with changes in f, b, and R.

1.2 COUPLING TO BLASTING CABLES

Equations (1) and (2) can be used to estimate the induced current in blasting cables. Equation (1) shows that the induced current in a blasting cable is dependent upon the conductor spacing, distance to the transmit loop, magnetic moment and frequency.

Assuming the worst case, the induced current becomes

$$I_L = \frac{\mu_o \; f \; M \; b}{R^2 \; 2(200\Omega)}$$

$$= \frac{(4\pi \times 10^{-7})(10^6)(4)(.04)}{(1)^2 \; 2(200)} \tag{18}$$

$$= 502 \text{ microamperes.}$$

It is believed that this level of induced line current is below the blasting threshold.

1.3 COUPLING TO AC POWER LINES AND OTHER SMALL CONDUCTOR SPACING STRUCTURES

The loop-line-loop signal level for b = 0.04 meters conductor spacing is shown below.

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Table E
Receiver Input Signal (V_R)

For each unit of magnetic moment b = .04 meter

Frequency kHz	Receiver Voltage uV
60	0.36
300	1.79
600	3.6

Above 300 kHz, AC power lines and other small conductor spacing structures
will provide a suitable transmission medium. As was the case with
trolley wire, passageway repeaters will increase the system performance.

1.4   FREE SPACE SPHERICAL PROPAGATION

Two geometries will be considered-the "aligned case" where only $\vec{H}_r$
(complex) exists and the "coplanar case" where only $\vec{H}_\theta$ exists.
Intermediate directions create elliptic polarization where both $\vec{H}_r$ and
$\vec{H}_\theta$ exist.   Then the magnitude of the major axis may be computed, which,
with suitable orientation of the receiving loop, will give the best emf
in the receiving loop.   This complication will not be considered in this
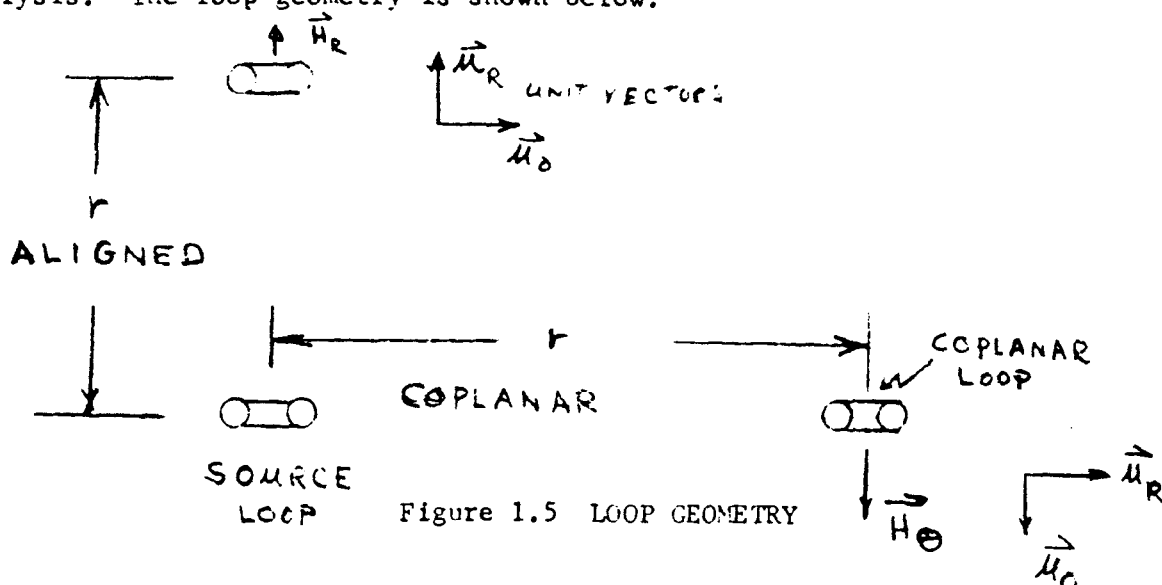analysis.   The loop geometry is shown below.

Figure 1.5 LOOP GEOMETRY

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The vector representation of the magnetic fields* are given by

$$\vec{H}_\theta = -\frac{2\eta^2 M}{\lambda^3}\left[\frac{1}{Br} - \frac{j}{(Br)^2} - \frac{1}{(Br)^3}\right] \quad \begin{array}{l}\theta = 90^o \\ \sin\theta = 1\end{array} \qquad (19)$$

and

$$\vec{H}_r = \frac{4\pi^2 M}{\lambda^3}\left[\frac{j}{(Br)^2} + \frac{1}{(Br)^3}\right] \quad \begin{array}{l}\theta = 0^o \\ \cos\theta = 1\end{array} \qquad (20)$$

*(These components are calculated in the receiving situation shown where their respective maxima occur.)

Values of the independent variables have been determined for the communication system and are shown in the following tables.

PARAMETERS

Frequency kHz	λ	B radians/m
60	5000m	1.257×10^{-3}
500	600m	10.472×10^{-3}
700	428.27m	14.66×10^{-3}

COEFFICIENTS OF FORMULA (uA/m)

Frequency kHz	$2\pi^2/\lambda^3$	$4\pi^2/\lambda^2$
60	1.58×10^{-3}	3.16×10^{-3}
500	0.091	0.182
700	0.25	0.5

FIELD COMPONENTS IN uA/m

r meters	Br radians	$\vec{H}_r$	$\vec{H}_\theta$	$|\vec{H}_r|$	$|\vec{H}_\theta|$
		(At 60 kHz)			
79.6	.1	3.16 + j .316	1.56 + j .158	3.16	1.57
7.96	.01	3160 + j 31.6	1580 + j 15.8	3160	1580
100	1.26	1.58 + j .199	0.78 + j 0.1	1.59	0.786
150	.188	0.471 + j .089	.233 + j .044	0.48	0.237
200	0.2514	0.198 + j .05	.093 + j .025	0.204	.096
500	0.63	0.0126 + j .008	.0077 + j .004	0.015	.0087

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r meters	Br radians	$\vec{Hr}$	$\vec{H_\theta}$	$|\vec{Hr}|$	$|\vec{H_\theta}|$
		(At 500 kHz)			
10	.105	157.2 + j 16.5	77.74 + j 8.25	158	78.1
100	1.05	.157 + j .165	-.008 + j .083	.227	.083
500	5.25	.001 + j .006	-.017 + j .03	.007	.017
		(At 700 kHz)			
100	1466	.159 + j .233	.09 + j .12	0.282	0.148

The voltage source in the receiving loop can be evaluated from

$$\mathcal{E} = . \; |\vec{H} \text{ applicable} | \mu_0 \omega . A . N \quad (21)$$

$$\underbrace{\frac{dB}{dt}}_{\frac{d\phi}{dt}}$$

where $\vec{H}$ applicable is the component of H which threads the loop in a direction normal to its plane (time magnitude).

By using Equations (10) and (21), the value of receiver terminal voltages have been computed and shown in the following table.

Table F
Receiver Input Signal (V_R)

For each unit of magnetic moment $R_L = 4.7$ ohms

Distance meters	Frequency kHz	Input Signal Voltage (uV) Aligned	Coplanar
500	60	0.016	--
150	60	0.518	0.015
100	60	1.717	0.08
100	500	2.04	0.74
100	700	3.558	1.876

The table indicates that loop-loop communications may exceed 100 meters. Further, (all system parameters being equal) the range increases with operating frequency.

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1.5 COMMUNICATION SYSTEM PERFORMANCE

By using prior U.S.B.M. mine communications research data, the operating range and fade margins along the signalling paths can be determined.

The maximum theoretical communication range (rm) at a given operating frequency can be determined by comparing the expected or measured signal level with either the measured or theoretical noise level.

The National Bureau of Standards measured noise data provides a basis for the determination of the noise floor in a communication system. The data applies in the "vicinity of several different locations in the mine." At other locations, the noise signals are attenuated by the appropriate propagation mechanism.

The 12 dB SINAD sensitivity (Hs in dB re 1 uA/m) characteristic of the Al receiver ($N=7$, $A=0.2$ m^2) has been included on the measured noise data (ADL HO 346045) shown on Figure 73. Figure 73 shows that in the vicinity of the noise sources, the mine generated noise will dominate the receiver signal processing networks. Assuming that transmission line signal propagation also exists in the vicinity, the expected signal level has also been included ($M=1$). The data shows that LLL communications will enjoy a favorable signal to noise ratio for operating frequencies of greater than 100 kHz. The data shows that a noise operated squelch receiver design would be troublesome in many mine locations. The average mine noise representation provided by ADL ($N \approx 34-20$ Log_{10} $(f/10)$) has been

Plotted Noise Levels Normalized to 12kHz Bandwidth — Measured with 1kHz
Instrumentation Bandwidth
— — — —   Vertical Field Component
——— · ———   Vertical Component — "Quiet Time"
————————   Horizontal Field Component

McElroy Mine — Continuous Miner Section and Nearby Rail Haulageway

- ○   Near end of rail haulage line
- -○-   Near intersection of rail hauiageway and conveyor belt
- ⅄   Near operating continuous mining machine
- ▯   Near section power distribution center

FIGURE 73   REPRESENTATIVE RMS MAGNETIC FIELD NOISE LEVELS
MEASURED IN THE McELROY COAL MINE
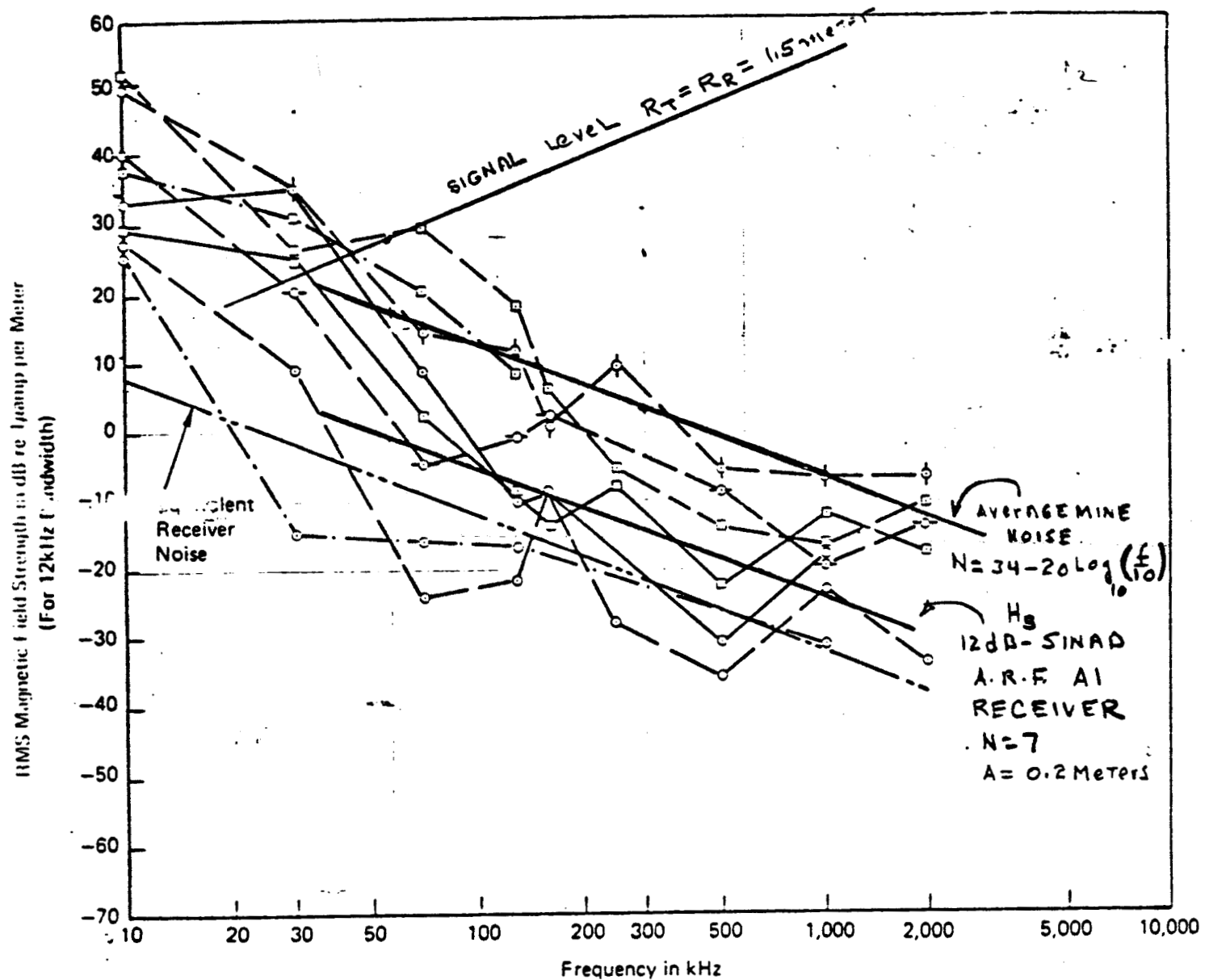
Source: National Bureau of Standards (Report NBSIR 74-389, June 1974)

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included to show that the McElroy noise data is about 10 dB less than the noise expected in an average mine. It should be noted that the noise signal exhibits a 1/f dependency (6 dB per octave change).

The signal levels appearing at other locations in the mine appear in the Collins report (IIO 366028).

Representative data from the Collins report has been included for range analysis purposes. By including the Al receiver sensitivity (Hs) point on the appropriate signal characteristic, the operating range can be determined (M=2.5). The appended Figure 26 shows that the range at 485 kHz will exceed 300 meters in the Robinson Run mine. Also, the Al receiver sensitivity may be better than the sensitivity of the "survey" receiver. That data may show that mine generated noise dominated the survey receiver. Figure 2-10 seems to indicate that the survey receiver has a slightly better sensitivity. The data also shows that the mine generated noise is quite low in the measuring location.

The measured signal data (T.Cory) indicates that the minimum operating range of a cellular repeater will exceed 100 meters in a conductor-free area. By coupling to trailing cables and other conductors, the expected communication range will cover the working areas of the mine.

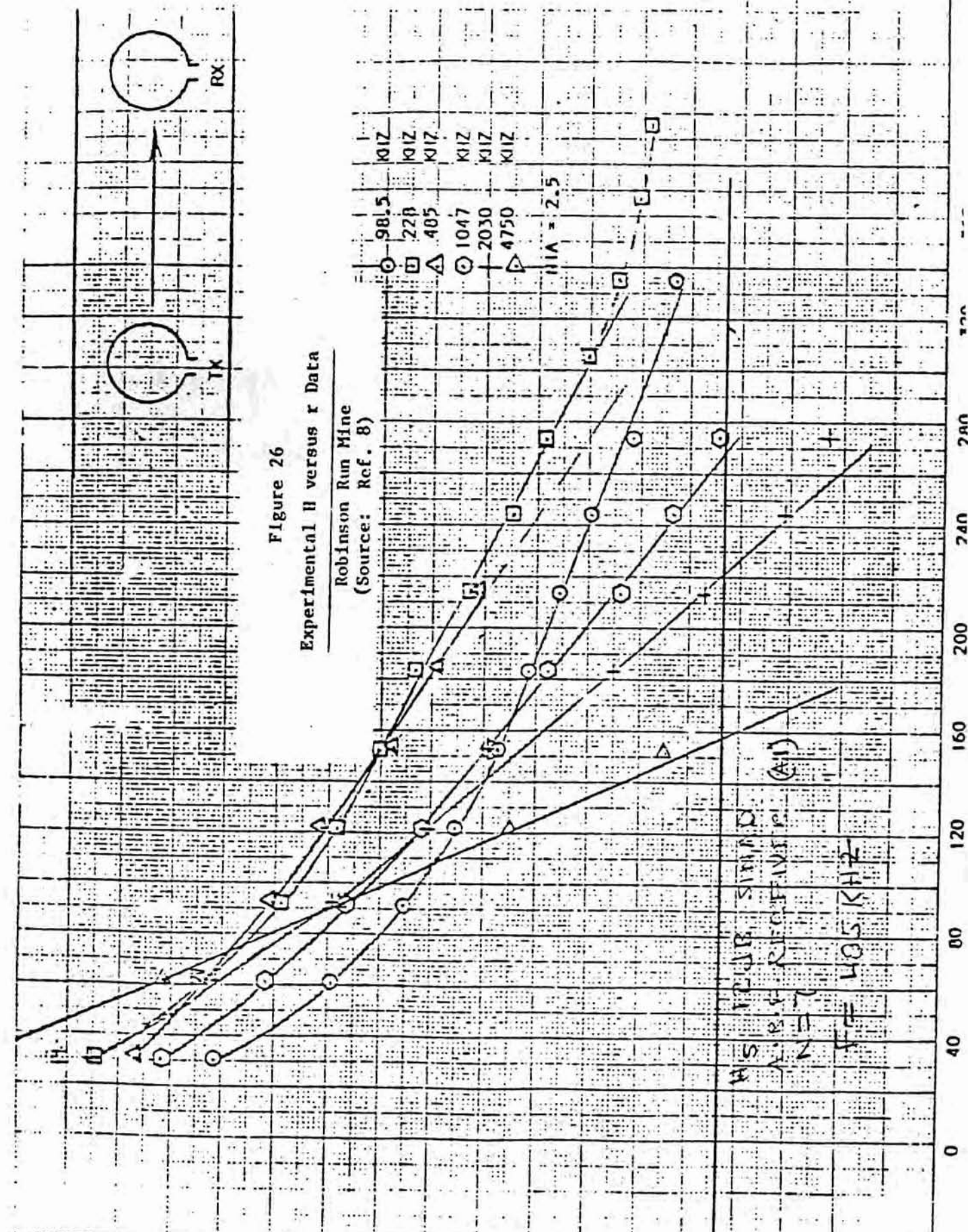Figure 26

Experimental II versus r Data
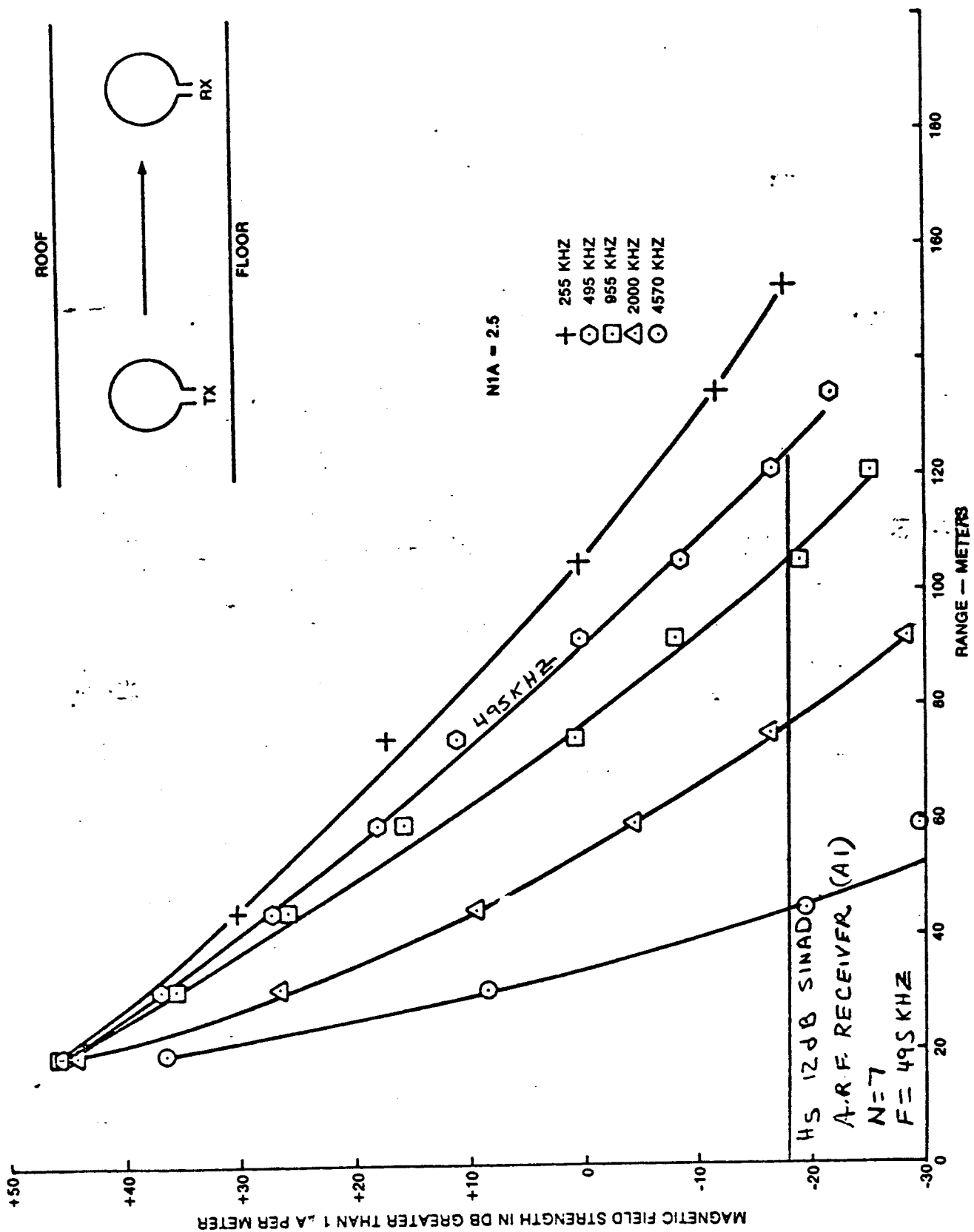
Robinson Run Mine
(Source: Ref. 8)

Figure 2-10. Magnetic Field Strength Vs Range in Quasi-Conductor-Free Area of 1 Main East for Coplanar IIMD Orientation of Transmit and Receive Antennas.

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1.6 MAGNETIC MOMENT

The magnetic moment is defined as (M)

$$M = NA\sqrt{\frac{P_o}{R_L}} \qquad (22)$$

where P_o is the transmitter output power in watts, and

R_L is the series resistance of a single tuned loop antenna.

The problem of maximizing the magnetic moment for a given transmitter output power (P_o) is carried out by increasing N and decreasing R_L. The geometry of the loop antenna (man-pack) has limited the loop area (A) to range between 0.2 and 0.25 meters^2. An N turn loop produces a loop inductance (L) of 80 microhenrys (ADL HO 346045 page 58). The series resistance of a single tuned loop antenna is

$$R_L = \frac{\omega L}{Q_{CKT}} \qquad (23)$$

where Q_{CKT} is the ratio of center frequency to 3 dB bandwidth (f_o/BW). In a single tuned loop antenna network

$$Q_{CKT} = Q_U \qquad (24)$$

where Q_U is the unloaded Q_U of the inductor (L).

Unloaded Q_U of greater than 80 can be achieved with an $N=8$ antenna inductor. At 60 kHz, the series resistance (R_L) is 0.38 ohms. Using equation (22), the magnetic moment (M) for a four (4) watt transmitter is

$$M = 8 \ (.2)\sqrt{\frac{4}{.38}}$$

$$= 5.19 \text{ ampere-meter}^2 \ .*$$

*Note: This maximum magnetic moment is based upon the realizable Q_U of the loop antenna structure.

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The 3 dB transmission bandwidth of the maximum magnetic moment loop is

$$BW = \frac{f_o}{Q_{CKT}} \qquad (24)$$

$$= \frac{60,000}{80} = 750 \text{ Hz.}$$

This loop antenna would not be useful because the transmission bandwidth is too narrow to accomodate the occupied bandwidth of the modulated FM signal (BW = 10,000 Hz).

The current design practice is to degrade the Q_U of the single tuned loop by artificially adding dissipation loss(resistance) to the loop circuit. The Q_U must be degraded to produce a 3 dB transmission bandwidth (Δ f) of 10,000 Hz.

$$Q_{CKT} = Q_U = \frac{f_o}{\Delta f}$$

$$= \frac{60,000 \text{ Hz}}{10,000 \text{ Hz}} = 6 \text{ (maximum).}$$

As a result of the FM occupied bandwidth constraint, R_L must be increased to 5.0 ohms thereby degrading M to

$$M = 8 \ (.2) \sqrt{\frac{4}{5}}$$

$$= 1.43 \text{ ampere-meter}^2.$$

The minimum intrinsically safe resistance ($\hat{R}_L$) for a 4 watt transmitter is

$$\hat{R}_L = \left(\frac{NA}{2.5}\right)^2 P_o \qquad (25)$$

$$= \left(\frac{(8)(.2)}{2.5}\right)^2 4 = 1.638 \text{ ohms.}$$

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The artificial increase in resistance (R_L) requires three times as much transmitter output power to achieve the maximum safe magnetic moment.
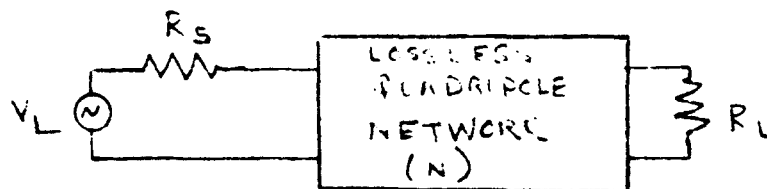
A feasible solution to the magnetic moment optimization problem is to use a second order (or higher order) loop antenna network. In a first order (series tuned) filter network

$$Q_U = Q_{CKT} = \frac{\omega L}{R_L} \qquad (26)$$

Therefore, there exists no possibility of maximizing M given BW, N and L (i.e. too many constraints).

In a higher order (n) lossless purely reactive quadripole network $(n \geq 2)$, the transmission bandwidth of the network is independent of the unloaded Q_U of the network elements.* As a result, there exists a feasible means of holding BW invariant over a reasonable range of R_L.

The higher order loop antenna structures can be designed using the insertion loss synthesis procedure of Darlington. In the synthesis process, an all pole lossless quadripole network is realized as shown below.

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* In practice, the transmission bandwidth is dependent on the unloaded Q_U.

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In the insertion loss method, power is sourced from the generator and delivered to the load plane resistance (R_L). The network has the property of an impedance matching network. The quadripole network is bilateral and useful as both a transmitting and receiving antenna structure.

Quadripole networks can be designed with different termination ratios ($r = R_S/R_L$). Later in the actual design of the second order loop antenna it will be obvious that

$$R_S = 1.63 \text{ ohms,}$$

$$R_L = 50 \text{ ohms, and}$$

$$r = 1.63/50 = 0.03 .$$

In the filter design problem the voltage source will represent the source voltage (V_L) in the tuned loop (see Figure 1.2). Thus part of the second order filter must have the usual loop topology.
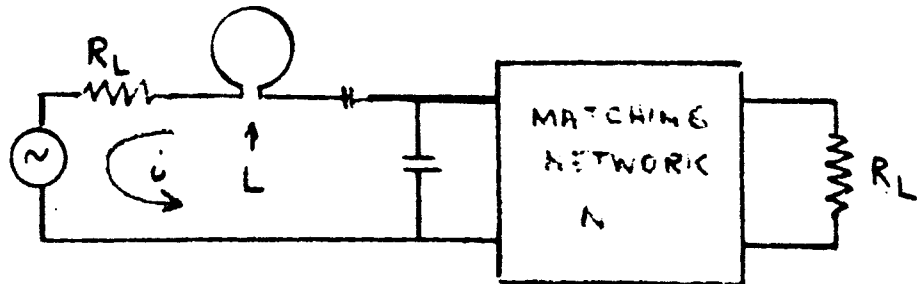
where the circulating loop current i flows in a resonant circuit. The energy stored in the inductor is given by

$$E = \tfrac{1}{2} L \, i_{pk}^{2} \quad \text{and} \qquad (27)$$

the energy dissipated per cycle is lost in R_L and given by

$$P_d = \frac{i^2 R_L}{2 \pi f} .$$

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The quality factor Q of this part of the network is

$$Q = \frac{E}{P_d}$$

$$= \frac{\tfrac{1}{2}L\, i^2}{i^2 R_L / 2\pi f} = \frac{\omega L}{R} = Q_U .$$

If the network topology can be realized as a second order passband filter,

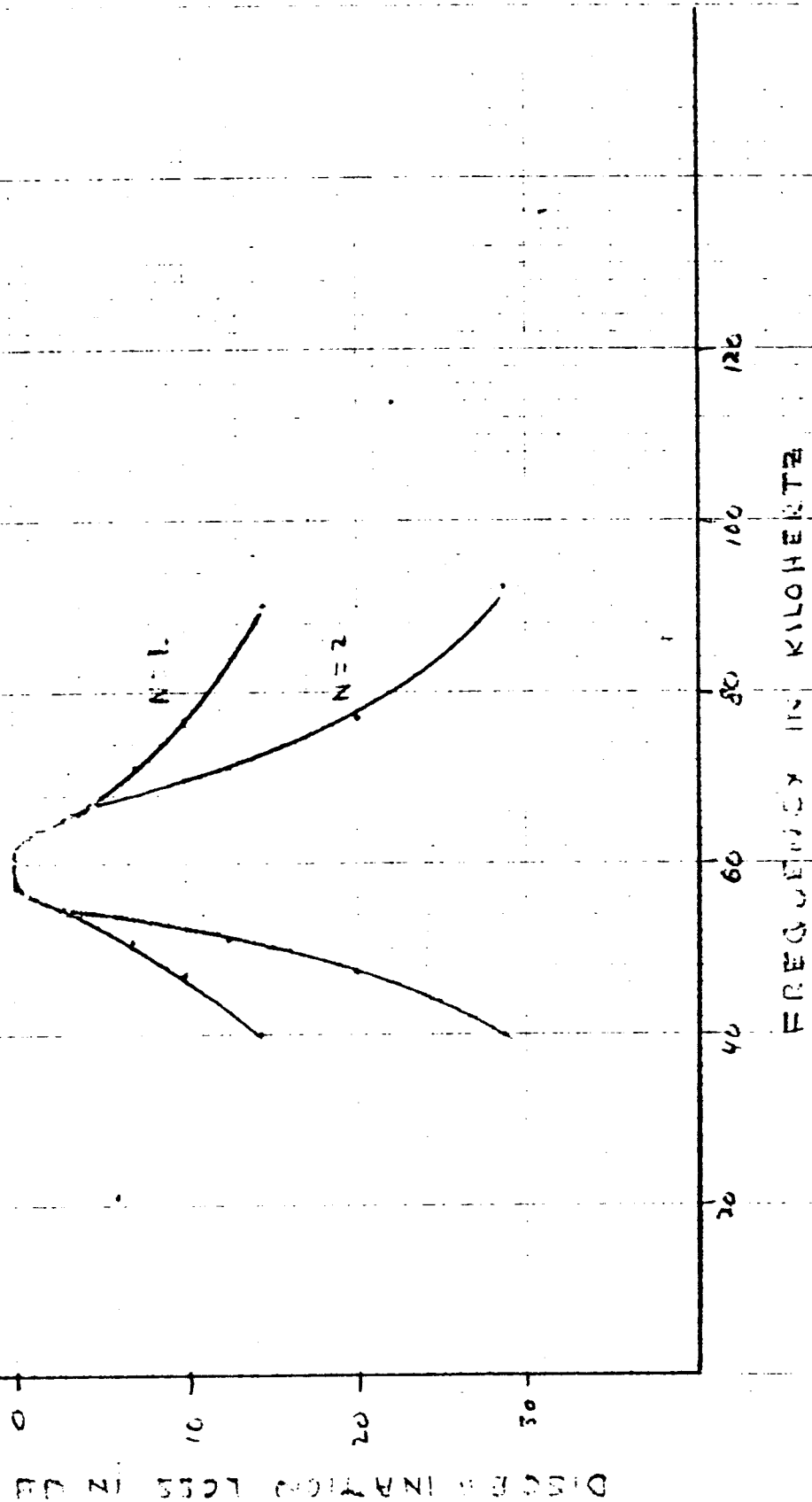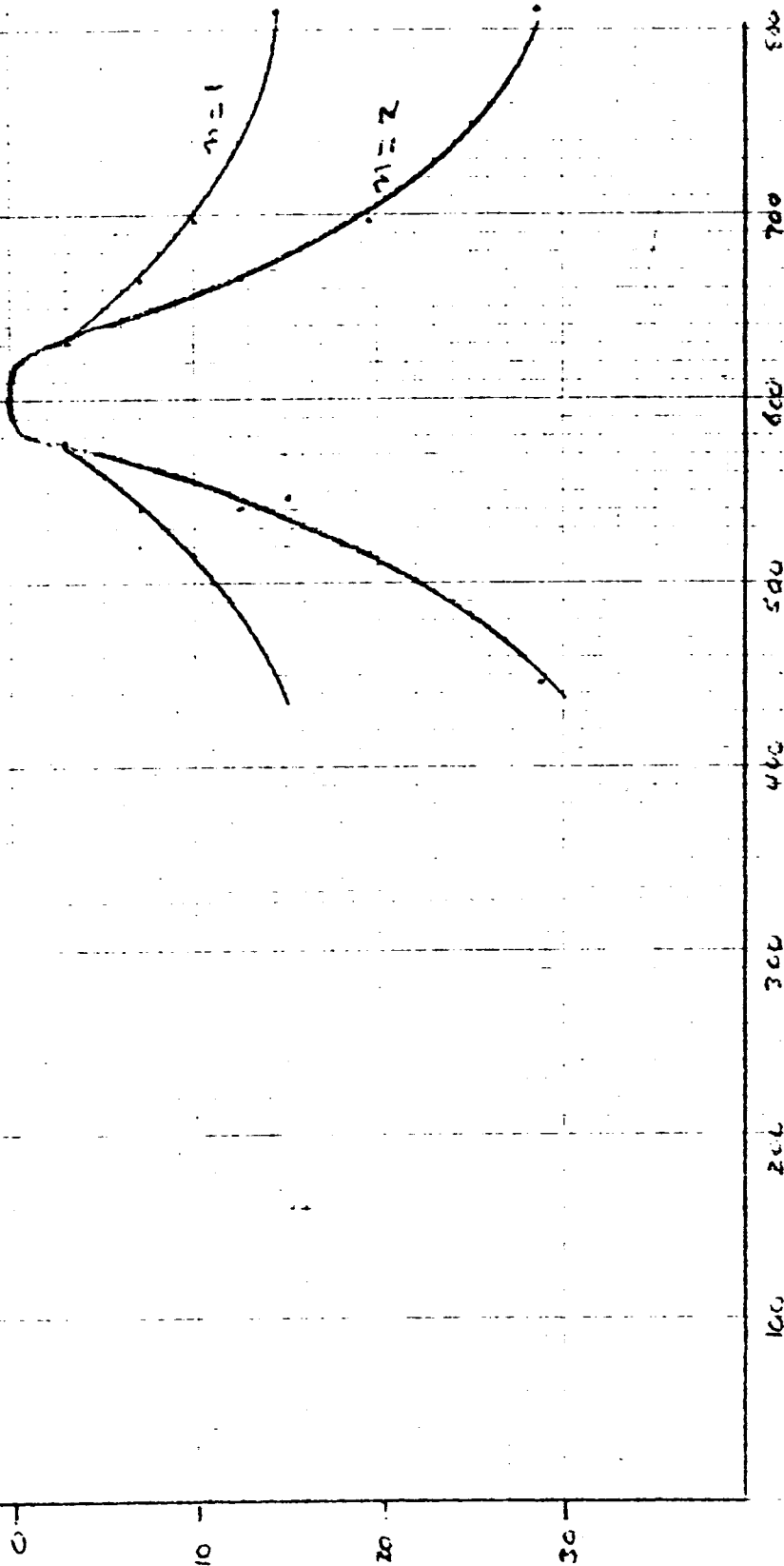the power transfer characteristic will be as shown in the following figure.

DISCRIMINATION LOSS IN dB

FIRST AND SECOND ORDER
LOOP ANTENNA NETWORK

$F_o = 60 KHz$

$BW = 10 KHz$

N = 1

N = 2

FREQUENCY IN KILOHERTZ

DISCRIMINATION LOSS IN dB

DISCRIMINATION LOSS IN dB

FIRST AND SECOND ORDER

LOOP ANTENNA NETWORKS

$f_c = 600$ KHz

$BW = 60$ KHz

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In the following section, a mesh coupled resonator filter will be used to design the second order antenna network.

One advantage of the second order loop antenna is that the transmission bandwidth can be as wide as 10 to 15 % of F_o.

In a second order tuned loop antenna, the loop tuning will not be critical as in the first order loop case. The antenna will also provide more discrimination loss against mine noise, and out of band signal.

Scanning transceivers with sufficiently close operating frequencies will not require automatic tuning antenna structures.

A disadvantage of this antenna is that it requires more circuit elements. When automatic tuning is required, a minimum of four circuit elements must be switched, maximizing the loop antenna transmit magnetic moment, also increasing the preformance of the receiving antenna.

TRANSMISSION BANDWIDTH RESTRICTION

MESH COUPLED RESONATOR
LOOP ANTENNA ($n=2$)

—— RECOMMENDED
– – – MAXIMUM

FREQUENCY IN KHz

TRANSMISSION BANDWITH IN KHz

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1.6.1   Design of Coupled Resonator Loop Antenna Networks

A coupled resonator antenna network can be designed from the low-pass prototype model with normalized element values (R_s, $\overline{L}_1$, $\overline{C}_2$,...R_L). The prototype network is shown below.
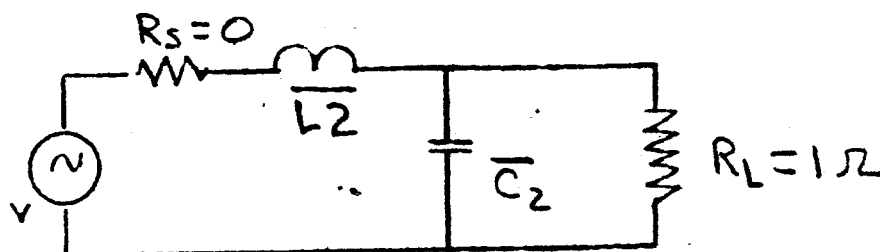

Figure 1.6   SECOND ORDER - LOW PASS PROTOTYPE NETWORK
(Special Case with R_s = 0)

The synthesized lowpass prototype element values are:

R_s = 0 ohms

$\overline{L}_1$ = 1.414

$\overline{C}_2$ = 0.707,   and

R_L = 1 .

This particular prototype network will match a voltage source (zero source resistance) to a given value of load resistance (RL).   The design approach assumes that loop antenna is operating as a receiver with an induced voltage source V (emf).

The transformation of the prototype circuit parameters to the mesh coupled resonator filter creates a pass band filter shown below.

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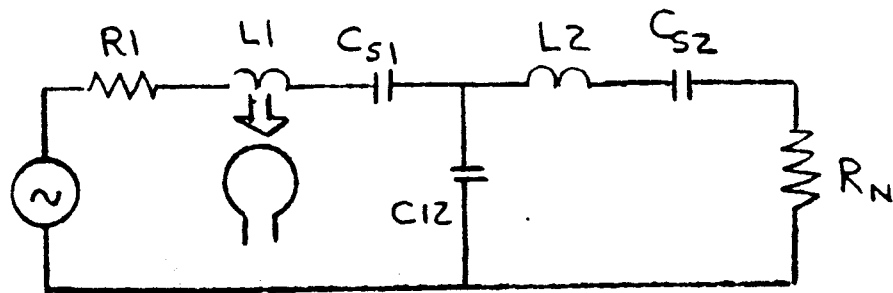TITLE   WIRELESS MF COMMUNICATION SYSTEM

Figure 1.7   MESH COUPLED RESONATOR FILTER NETWORK

The mesh coupling capacitor is C12.

Coupled resonators are popular in filter design because the actual filter element values can be chosen in the design process. By way of contrast, in the synthesis of the low pass prototype, the element values are the result of a continued fraction expansion; hence, are not selectable.

The principal design equations for the mesh capacitor coupled structure are shown in the table below.

Table G
Mesh Capacitor Design Equations

$$C_{12} = \frac{1}{\omega_0 \, \Delta\omega \, \sqrt{L_1 L_2} \, k_{12}}$$

$$k_{12} = \frac{1}{\sqrt{L_2 C_1}}$$

$\Delta\omega =$ 3 dB bandwidth

$L_1 =$ antenna inductance (80 ull)

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The value of C_{s2} is

$$C_T = \frac{1}{\omega_0^2 L_2} = .125 \text{ uf}$$

then

$$C_{s2} = 0.156 \text{ uf}$$
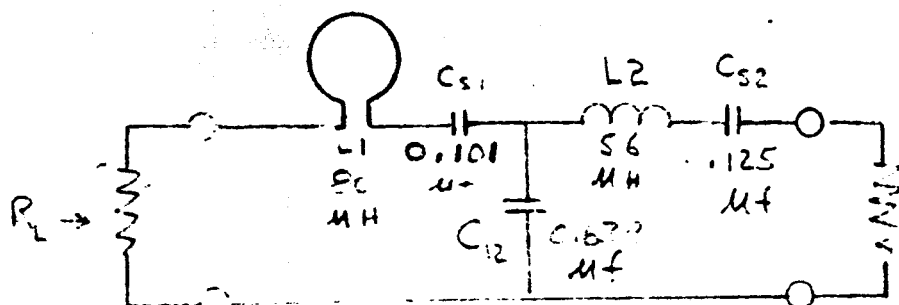
The antenna network is shown in Figure 1.8.

Figure 1.8 MESH COUPLED ANTENNA NETWORK
(BW = 10,000 Hz, f_o = 60,000 Hz)

The series resistance of the tuned loop circuit is R_L.

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APPENDIX A

Table A-1
Loop Antenna Parameters

Frequency in kHz	Q for BW	Turns	L uH	R ohms
60	6	8	80	5
300	30	7	75	4.7
500	50	7	75	4.7
700	70	7	75	4.7

Assumption:

Loop area (bandolier) $.25m \times .8m = .2m^2$

$$L \approx 2.5 A^{\frac{1}{2}} N^2 \text{ (uH)}$$

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APPENDIX B

DISCRIMINATION LOSS AND BANDEDGE FREQUENCIES
FOR A
FIRST AND SECOND ORDER TUNED LOOP ANTENNA

Table B-1

Center Frequency (f_o) 60 kHz
Bandwidth (BW_{CKT}) 10 kHz

Bandwidth	Bandedge Frequency in kHz		Discrimination loss	
BW_D	f_L	f_H	n=1	n=2
BW_{CKT}	55207	65207	3	3
$2BW_{CKT}$	50827	70827	6.9	12.
$3BW_{CKT}$	46846	76846	9.9	19.
$5BW_{CKT}$	40,000	90,000	14.5	28.

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Table B-2

DISCRIMINATION LOSS AND BANDEDGE FREQUENCIES
FOR A
FIRST AND SECOND ORDER TUNED LOOP ANTENNA

Center Frequency (f_c)   600 kHz
Bandwidth (BW_{CKT})   60 kHz

Bandwidth	Bandedge Frequency in kHz		Discrimination loss	
	f_L	f_H	$n=1$	$n=2$
BW_{CKT}	570750	630749	3	3
$2BW_{CKT}$	542992	662992	6.9	12.
$3BW_{CKT}$	516712	696712	9.9	19.
$5BW_{CKT}$	438585	820820	14.5	28.